

1 PILOTLESS, WIRELESS, TELECOMMUNICATIONS APPARATUS, SYSTEMS
2 AND METHODS HAVING EQUALIZATION
3

4 This is a divisional of U.S. Serial No. 10/628,943
5 filed July 29, 2003.

6 BACKGROUND OF THE INVENTION
7

8 1. Field of the Invention

9 The present invention relates to telecommunications.
10 The invention more particularly relates to wireless
11 telecommunications apparatus, systems and methods which
12 implement data transmission via a plurality of
13 telecommunication channels such as radio channels with
14 variable parameters. More specifically, the invention
15 relates to wireless systems with multicarrier transmission,
16 although it is not limited thereto.

17

18 2. State of the Art

19 In wireless data transmission systems, a signal is
20 subjected to several frequency conversions with respective
21 shifting of its carrier frequency and initial phase. In
22 mobile systems, the carrier frequency is additionally
23 subjected to the Doppler effect. In addition, the signal
24 phase at the receiving point depends on the time interval

1 of radio signal propagation in the communication channel,
2 and this time interval is changed because of both the
3 change of the signal propagation path and the change of
4 properties and parameters of the propagation media. In
5 wireless multipath channels, the change of any single
6 interference component (its amplitude or/and phase) causes
7 the change of the received signal phase as a whole. As a
8 result, the initial signal phase has a constant component
9 and a varying, typically slowly changing component.
10 Usually, in wireless systems, the constant component is
11 compensated in the receiver during the preamble by
12 estimating frequency offset and frequency equalizer
13 adjustment utilizing a special pilot signal.

14
15 Optimal signal processing in data transmission systems
16 and wireless telecommunication systems is based on certain
17 *a priori* information about received signals and channel
18 characteristics. This information includes symbol time
19 interval, carrier initial phase, signal attenuation,
20 signal-to-noise ratio and other service parameters, which
21 are extracted from the received signal by means of special
22 functions such as clock synchronization, carrier recovery,
23 signal equalization, channel estimation, etc. In channels
24 with variable characteristics, such as multipath wireless

1 channels, the above-mentioned service parameters change
2 over time, and their estimation, in order to remain
3 current, requires special adaptive or tracking procedures.

4
5 Typically in wireless systems, service parameter
6 estimation and tracking are based on utilization of special
7 pilot signals. Two types of pilot signals are usually
8 used: preamble pilots transmitted during a preamble before
9 data transmission, and accompanying pilots transmitted
10 during the whole communication session in parallel with
11 data transmission. As a rule, these two types of pilots
12 have not only different parameters but also provide
13 different functions.

14
15 The preamble pilot consists of few symbols and takes a
16 comparatively small part of the communication session. It
17 is used for automatic gain control (AGC), clock
18 synchronization, initial frequency offset correction,
19 preliminary carrier phase adjustment, as well as for
20 channel parameters estimation. For example, in a WLAN
21 system according to the IEEE802.11a standard, the preamble
22 pilot contains two training sequences: a short training
23 sequence, and a long training sequence. The short training
24 sequence consists of ten short OFDM symbols with duration

1 0.8 μ s, and the long training sequence consists of two long
2 OFDM symbols with duration 3.2 μ s. Each short OFDM symbol
3 is a sum of twelve phase-modulated carriers with numbers:
4 2, 6, 10, 14, 18, 22, 26, 30, 34, 38, 42, 46, 50. Each
5 long OFDM symbol is a sum of all fifty-two phase modulated
6 carriers. The short and long training sequences are
7 separated by a guard interval with a duration of 1.6 μ s.
8 The total duration of the preamble pilot signal (training
9 signal) is 16 μ s, which is 80% of a whole service signal,
10 transmitted before data, but it is a very small part of the
11 communication session as a whole.

12

13 The IEEE standard specifies that the short training
14 sequence should be "used for AGC convergence, diversity
15 selection, timing acquisition, and coarse frequency
16 acquisition in the receiver", and the long training
17 sequence should be "used for channel estimation and fine
18 frequency acquisition in the receiver" (Section 17.3.2.1).
19 So, the preamble pilot, as a rule, does not considerably
20 decrease the average data rate of the system (system
21 capacity), and this type of pilot signal is not the focus
22 of this invention.

23

1 In contrast to the preamble pilot signal, the
2 accompanying pilot signals are usually transmitted during
3 the whole communication session in parallel with data
4 transmission. The accompanying pilot signals are typically
5 used for adaptive equalization, for frequency offset
6 tracking, and for current adjustment of carrier phases to
7 provide improved coherent signal processing. For example,
8 in the WLAN system according to the IEEE802.11a standard,
9 the accompanying pilot signal consists of four pseudo-
10 randomly modulated carriers. The standard specifies: "In
11 each OFDM symbol, four of the carriers are dedicated to
12 pilot signals in order to make coherent detection robust
13 against frequency offset and phase noise. These pilot
14 signals shall be put in carriers -21, -7, 7, 21. The
15 pilots shall be BPSK modulated by a pseudo binary sequence
16 to prevent the generation of spectral lines" (Section
17 17.3.5.8). So, in the OFDM WLAN system forty-eight
18 carriers are used for data transmission and four carriers
19 are dedicated to pilot signals; i.e., about 8% of the
20 system capacity, as well as transmitter power, is used for
21 pilot signal transmission.

22

23 Approximately the same portion of the system capacity
24 is wasted in the fixed wireless broadband systems according

1 to the IEEE802.16 standard (Section 8.3.5.3.4), in which
2 one constant pilot carrier is used per twelve data
3 carriers.

4
5 It should be noted that a decreasing real data rate is
6 not the only disadvantage of pilot utilization. When using
7 frequency spaced (i.e., frequency-separated) pilots for
8 phase adjustment of the carrier signals, the accuracy of
9 the phase adjustment is not sufficient for perfect coherent
10 processing, especially in multipath wireless channels. As
11 a matter of fact, the phases of the frequency spaced
12 carriers are not 100% correlated. Therefore, even if the
13 estimation of a pilot phase is perfect, the estimation of
14 an adjacent carrier phase may be not correct. Taking into
15 account this fundamental disadvantage of pilot systems, the
16 authors of the IEEE802.16 standard have proposed to use
17 variable location pilot carriers in addition to the
18 constant location pilot carriers. Variable pilots shift
19 their location each symbol with a cyclic appearance. This
20 technique allows a receiver to improve phase tracking
21 accuracy, but it leads to complicated synchronization and
22 additional capacity loss.

23

1 It should also be noted that existing approaches to
2 pilotless phase tracking system design are based on carrier
3 recovery techniques. See, J.Proakis, "Digital
4 Communications", 4th edition, McGraw-Hill, 2001, Section
5 6.2. Carrier recovery techniques provide individual phase
6 tracking for each carrier. They provide simple and
7 efficient solution for single carrier systems with small-
8 size constellations, but they are practically unacceptable
9 for multicarrier systems with multipoint QAM
10 constellations.

11

12 SUMMARY OF THE INVENTION

13

14 It is therefore an object of the invention to provide
15 apparatus, systems and methods which implement pilotless
16 telecommunications.

17

18 It is another object of the invention to provide
19 pilotless telecommunication systems which provide desired
20 receiving functions.

21

22 It is a further object of the invention to provide
23 pilotless telecommunications systems which extract
24 information from signal-bearing data in order to conduct

1 one or more of adaptive equalization, frequency offset
2 tracking, and current adjustment of carrier phases to
3 provide improved coherent signal processing.

4

5 It is an additional object of the invention to provide
6 pilotless telecommunication systems which transmit data
7 without any pilot signals and can therefore use all system
8 bandwidth exclusively for data transmission, while still
9 providing all receiving functions based on extraction of
10 all necessary information from signal-bearing data.

11

12 Another object of the invention is to provide general
13 methods and apparatus for pilotless frequency offset
14 compensation and carrier phase tracking necessary for
15 optimal coherent processing of the received signals in
16 single-carrier and multi-carrier systems with different
17 modulation techniques, including any type of QAM
18 constellations.

19

20 A further object of the invention is to provide
21 simplified methods and apparatus for pilotless frequency
22 offset compensation and carrier phase tracking in
23 multicarrier systems with correlated between-carrier
24 phases.

1 An additional object of the invention is to provide
2 methods and apparatus for pilotless adaptive per-carrier
3 equalization in multicarrier systems.

4
5 Yet another object of the invention is to provide
6 pilotless signal equalization, frequency offset
7 compensation, as well as carrier phase tracking based on
8 algorithms which do not require complex signal processing
9 and can be implemented utilizing the existing demodulation
10 and decoding apparatus.

11
12 In accord with the objects of the invention, the
13 present invention broadly provides systems, methods and
14 apparatus which transmit signal-bearing data without
15 accompanying pilot signals and which provide receiving
16 functions based on extraction of information from the
17 signal-bearing data. Among these functions are frequency
18 offset compensation and carrier phase tracking.

19
20 According to one embodiment of the invention, an
21 optimal (in terms of minimum variance of phase estimates)
22 algorithm of phase adjustment is implemented in a pilotless
23 system, method, and apparatus by reducing and averaging
24 differential quadrature components of the received signal.

1 A "differential quadrature component" is defined as the
2 difference between the corresponding quadrature components
3 of a received signal and a decision signal. "Reduction" of
4 differential quadrature components of the received signal
5 consists of a linear transformation of the received signal
6 to the likely differential components of a reference
7 signal, which may be any predetermined vector. Averaging
8 of differential components of the reference signal provides
9 nonbiased and efficient estimates of the phase shift,
10 particularly if all decisions are correct.

11

12 It should be noted that differential components of the
13 received signal may be used for optimal soft decision
14 decoding as well as for mode assignment and adaptation to
15 channel conditions as disclosed in co-owned U.S. Serial No.
16 10/342,519 entitled "Methods, Apparatus, and Systems
17 Employing Soft Decision Decoding", and U.S. Serial No.
18 10/406,776 entitled "Mode Adaptation in Wireless Systems",
19 both of which are hereby incorporated by reference herein
20 in their entireties. In the present invention, the
21 differential components are utilized for estimation of
22 frequency offset and carrier phase shift.

23

1 According to an alternative embodiment of the
2 invention, phase adjustment may be accomplished via
3 reduction and averaging of quadrature components of the
4 received signal. It should be appreciated that in either
5 embodiment (i.e., phase adjustment utilizing reduction and
6 averaging of differential quadrature components, or phase
7 adjustment utilizing reduction and averaging of quadrature
8 components), a demapping procedure is accomplished with
9 linear operations and without direct calculation of the
10 carrier phase. This is in contrast to the prior art
11 approach which finally calculates the phase of the received
12 carrier for the proper correction of the reference signals.
13 See, e.g., J.Proakis, "Digital Communications", 4th
14 edition, McGraw-Hill, 2001, Section 6.2.

15
16 According to a further aspect of the invention, based
17 on estimates of differential quadrature components or
18 quadrature components of the reference signal, two
19 embodiments are provided for the demapping procedure within
20 the phase tracking loop. A first embodiment corrects the
21 received signal, while a second embodiment corrects the
22 constellation points.

23

1 The first embodiment, which, in most circumstances is
2 the desirable one from the implementation point of view,
3 includes the proper rotation of the received signal
4 (correction of the received coordinates) with further
5 decision-making based on the corrected received signal
6 without changing constellation points. The advantage of
7 this method is that it does not need any correction of the
8 constellation points, and, as a result, preserves the
9 simplest decision-making procedure, based on a comparison
10 of the received coordinates with a limited number of
11 thresholds.

12

13 The second embodiment of implementing demapping within
14 the phase tracking loop, is based on estimates of
15 differential quadrature components or quadrature components
16 of the reference signal, and includes the proper rotation
17 of the constellation points (correction of the
18 constellation point coordinates) with further decision-
19 making based on the corrected constellation points. The
20 advantage of the second mechanism is that it provides
21 optimal adaptive processing without any changing of the
22 received signal. In other words, the receiver does not
23 spend processing time for transformation of each received
24 symbol, and all processing relates only to constellation

1 point correction. The advantage is considerable primarily
2 for small size constellations, for example, for QPSK
3 modulation techniques.

4

5 According to another aspect of the invention,
6 algorithms are provided which implement a general method of
7 phase shift estimation in single carrier and multicarrier
8 pilotless wireless systems with uncorrelated between-
9 carriers phase shifts. In the multicarrier case, they can
10 provide individual phase tracking for each carrier.

11

12 According to other aspects of the invention, special
13 simplified algorithms of frequency offset compensation and
14 phase shift tracking for multicarrier systems with
15 correlated between-carrier phases are provided. The
16 simplifications are based, first, on replacing averaging in
17 the time domain with averaging in frequency domain, and,
18 second, on the utilization of the same phase shift estimate
19 for all carriers. As with the general algorithms, the
20 final demapping procedure in the simplified algorithms may
21 use either correction of the received signal or correction
22 of the constellation points.

23

1 According to yet another aspect of the invention, a
2 further simplification of the pilotless multicarrier
3 system, apparatus, and method is possible when carrier
4 phase shifts are correlated and comparatively small. For
5 this particular case, an extremely simplified algorithm for
6 phase tracking is provided which is based on the estimation
7 of only one differential component of the simplest
8 reference vector. In one embodiment related to this aspect
9 of the invention, the phase shift is efficiently corrected
10 by majority-type algorithms which are based on an
11 accumulation of differential component signs. The simplest
12 version of the majority-type algorithms provides changing
13 carrier phases with a constant small increment. In this
14 case the phase adjustment algorithm determines only a
15 direction of the adjustment which is provided by the proper
16 majority vote procedure.

17

18 According to even another aspect of the invention, the
19 proposed methods, systems, and apparatus for carrier phase
20 tracking, which utilize estimates of differential
21 quadrature components or quadrature components of the
22 reference signal, can be further used for adaptive
23 equalization of the received multicarrier signals. In this
24 case, a per-carrier adaptive equalizer for multicarrier

1 wireless systems is provided and is based on estimates of
2 differential quadrature components of the reference vector.
3 The equalizer combines static and dynamic equalization
4 functions into a one-step adaptive procedure.
5

6 Additional objects and advantages of the invention
7 will become apparent to those skilled in the art upon
8 reference to the detailed description taken in conjunction
9 with the provided figures.
10

11 BRIEF DESCRIPTION OF THE DRAWINGS

12

13 Fig. 1 is a plot showing a signal constellation and
14 various vectors useful in understanding the invention.
15

16 Fig. 2 is a two-dimensional plot showing results of
17 stochastic simulation of random signal reduction and
18 averaging for 16-QAM constellation points, phase shifted by
19 $\pi/11$, in the AWGN channel.
20

21 Fig. 3 is a two-dimensional plot showing results of
22 stochastic simulation of random signal reduction and
23 averaging for 16-QAM constellation points, phase shifted by

1 $\pi/16$, in the AWGN channel, when all decisions are correct
2 ones.

3

4 Fig. 4 is a two-dimensional plot showing results of
5 stochastic simulation of random signal reduction and
6 averaging for the same conditions as in Fig. 3, when the
7 decisions have errors with symbol error rate 0.01.

8

9 Fig. 5 is a flow chart illustrating correction of the
10 received signals based on reduction and averaging of
11 differential quadrature components of the received signals.

12

13 Fig. 6 is a flow chart illustrating correction of the
14 constellation point coordinates based on reduction and
15 averaging of differential quadrature components of the
16 received signals.

17

18 Fig. 7 is a flow chart illustrating correction of the
19 received carriers in a multicarrier system with correlated
20 phase shift, based on differential quadrature components of
21 the received carriers.

22

23 Fig. 8 is a flow chart illustrating correction of the
24 constellation point coordinates in a multicarrier system

1 with correlated phase shift, based on differential
2 quadrature components of the received carriers.

3

4 Fig. 9 is a plot showing a constellation and various
5 vectors useful in understanding a simplified algorithm of
6 phase correction according to the invention.

7

8 Fig. 10 is a flow chart illustrating simplified
9 carrier phase correction in multicarrier systems based on
10 the differential quadrature components.

11

12 Fig. 11 is a flow chart illustrating the majority
13 algorithm of carrier phase correction in multicarrier
14 systems based on the differential quadrature components.

15

16 Fig. 12 is a flow chart for a per-carrier adaptive
17 equalizer, based on estimates of differential quadrature
18 components.

19

20 DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

21

22 According to one embodiment of the invention, an
23 optimal (in terms of minimum variance of phase estimates)
24 algorithm of phase adjustment in pilotless systems utilizes

1 reduction and averaging of differential quadrature
2 components of the received signal, where, as set forth
3 above, a "differential quadrature component" is defined as
4 the difference between the corresponding quadrature
5 components of a received signal and a decision signal.

6
7 According to an alternative embodiment of the
8 invention, phase adjustment in a pilotless system is
9 accomplished via reduction and averaging of quadrature
10 components of the received signal. Both embodiments
11 provide demapping using linear operations and without
12 direct calculation of the carrier phase. In addition, both
13 embodiments solve two major problems of pilotless systems:
14 the problem of fine phase adjustment, and the problem of
15 channel estimation.

16
17 Channel estimation includes two basic procedures:
18 channel quality estimation and channel parameters
19 estimation. The channel quality estimation is usually
20 based on signal-to-noise ratio (SNR) and/or on some
21 functions of the SNR, and it is used for mode assignment,
22 for adaptation to channel conditions, as well as for
23 optimal soft decision decoding. The channel quality
24 estimation algorithms and the corresponding apparatus and

1 systems based on the calculation of differential components
2 of the received signal are described in previously
3 incorporated U.S. Serial Nos. 10/342,519 and 10/406,776.

4
5 The channel parameters estimation is typically based
6 on channel pulse response or channel frequency
7 characteristics. In the case of multicarrier systems, for
8 example OFDM, a set of carrier amplitudes and initial
9 phases completely determine channel parameters necessary
10 for frequency equalization of the received signal. As is
11 described below, the methods of the invention for carrier
12 phase adjustment in pilotless systems which are based on
13 reduction and averaging of differential quadrature
14 components of the received signal, provide simultaneously
15 information applicable to channel parameters estimation in
16 terms of amplitudes and phases of frequency carriers. A
17 per-carrier equalizer for multicarrier wireless systems,
18 based on estimates of differential quadrature components of
19 the reference vector is provided. The equalizer combines
20 static and dynamic equalization functions into a one-step
21 adaptive procedure.

22
23 The methods, apparatus, and systems of the invention
24 provide carrier phase correction for both single carrier

1 and multicarrier wireless systems. The methods, apparatus,
2 and systems can be divided into two classes. The first
3 class includes general algorithms providing phase shift
4 compensation in pilotless wireless systems with
5 uncorrelated between-carriers phase shifts. The algorithms
6 are applicable for both single carrier and multicarrier
7 systems, including multicarrier systems with uncorrelated
8 carrier phases. The second class includes special
9 algorithms of phase shift compensation in multicarrier
10 systems with correlated carrier phases. The Wi-Fi
11 IEEE802.11a standard provides a typical example of a system
12 in the second class.

13

14 Before turning to Figure 1, it is useful to define
15 designations which are used in the algorithms of the
16 invention:

17 i - index of the current received symbol (vector)
18 within the sequence of received symbols;

19 n - index of the constellation points; $n=1,2,...m$;

20 X_i, Y_i - coordinates (real and imaginary components) of
21 the i 'th received vector (after equalization during the
22 preamble interval - static equalization);

1 X_{ir}, Y_{ir} - coordinates (real and imaginary components) of
 2 the i 'th reduced vector (result of reduction of coordinates
 3 X_i and Y_i);

4 dX_i, dY_i - coordinates (real and imaginary components)
 5 of the i 'th received differential vector;

6 dX_{ir}, dY_{ir} - coordinates (real and imaginary components)
 7 of the i 'th reduced differential vector;

8 X_0, Y_0 - coordinates (real and imaginary components) of
 9 the reference vector;

10 X_{cn}, Y_{cn} - coordinates (real and imaginary components) of
 11 the n 'th constellation points;

12 Δ_i - phase difference between the i 'th decision vector
 13 and the reference vector;

14 θ_n - phase difference between the reference vector and
 15 the n 'th constellation point;

16 A_n - amplitude of the n 'th constellation point;

17 a_i - amplitude of the i 'th decision vector;

18 A_0 - amplitude of the reference vector;

19

20 Fig. 1 shows a 16-point constellation in 2-dimensional
 21 space (x,y) , with the constellation points indicated by
 22 small crosses with corresponding binary combinations. Fig.
 23 1 also illustrates various vectors and variables. In Fig.
 24 1, a reference vector ("Reference vector") is shown having

1 coordinates (3,3) and being provided with a binary
 2 combination 1010. It should be noted at the outset, that
 3 conceptually, any vector in (x,y)-space may be considered
 4 as the reference vector. In practice, however, some
 5 choices may be more convenient than others. In particular,
 6 it is convenient when the reference vector coincides with
 7 the X-axis or Y-axis, such as e.g., vector (1,0) or vector
 8 (0,1).

9
 10 As seen in Fig. 1, it may be assumed that a signal
 11 with coordinates (X_i, Y_i) has been received (the "received
 12 vector"). Then a decision is made as to which
 13 constellation point is nearest the received vector. In
 14 Fig. 1, a decision is made that point 0010 with coordinates
 15 $(-3,3)$ is the nearest constellation point relative to the
 16 received vector; and thus a "decision vector" is shown in
 17 Fig. 1. Mathematically, the decision-making procedure is
 18 described as finding a minimum distance between the
 19 received signal and various constellation points:

20

$$21 \quad (X_{di}, Y_{di}) \Leftrightarrow \min_n ((X_i - X_{cn})^2 + (Y_i - Y_{cn})^2); \quad (1)$$

22

23 where (X_{di}, Y_{di}) are the coordinates of the decision,

1 (X_{cn}, Y_{cn}) are the coordinates of the n 'th constellation
2 point; $n=1,2, \dots, m$, and m is the number of constellation
3 points (constellation size). According to relationship (1)
4 above, the decision (X_{di}, Y_{di}) is a constellation point
5 providing a minimum value to the expression on the right
6 side.

7
8 It should be noted that each received vector contains
9 the proper information about a probable phase shift.
10 However, the received vector has an unknown phase and
11 amplitude due to information content. If a correct
12 decision regarding the transmitted vector is accomplished,
13 the unknown phase and amplitude can be removed via
14 rotation. The resulting vector is called the "reduced
15 vector" as it is shown in Fig. 1. As will be discussed
16 below, this transformation (or reduction) allows proper
17 parameters averaging.

18
19 As can be seen from Fig. 1, the reduced vector is
20 determined by means of simple rotation of the received
21 vector by angle Δ_i , which is equal to phase difference
22 between the decision vector and the reference vector.

23

1 A first embodiment of the invention is based on a
2 utilization of a "differential received vector", which is
3 equal to a difference between the received vector and the
4 decision vector in Fig. 1. As can be seen from Fig. 1, the
5 differential received vector begins from the origin, i.e.,
6 the (0,0) point of the (x,y)-space, and it is determined by
7 differential quadrature components, which are differences
8 between the corresponding quadrature components of the
9 received vector and the decision vector.

10

11 Calculation of the differential components of the
12 differential received vector is a part of a decision-making
13 procedure well known in the art (see, e.g., IEEE 802.11a,
14 Wireless LAN Medium Access Control (MAC) and Physical Layer
15 (PHY) specifications in the 5 GHz Band, Sections 17.3.2.1
16 and 17.3.5.8), and they are also calculated in well-known
17 soft decision decoding algorithms, as well as in the new
18 mode-adaptation methods described in the previously
19 incorporated patent applications. In other words, the
20 differential components are generally available as a
21 byproduct of different computations necessary for modem
22 functioning.

23

1 The differential quadrature components of the received
2 signal are:

3

$$4 \quad dX_i = (X_i - X_{di}) , \quad (2a)$$

$$5 \quad dY_i = (Y_i - Y_{di}) , \quad (2b)$$

6

7 where (X_{di}, Y_{di}) is the i 'th decision vector, which is
8 typically equal to the constellation point nearest to the
9 received vector (X_i, Y_i) .

10

11 The reduced differential vector (see Fig. 1) is
12 determined by means of simple rotation of the differential
13 received vector through angle Δ_i (which in turn, as
14 described above, is equal to phase difference between the
15 decision vector and the reference vector).

16

17 In the general case, transformation of the
18 differential received vector into the reduced differential
19 vector may be described as follows:

20

$$21 \quad dX_{ir} = (A_0/a_i)(dX_i \cos \Delta_i - dY_i \sin \Delta_i), \quad (3a)$$

$$22 \quad dY_{ir} = (A_0/a_i)(dY_i \cos \Delta_i + dX_i \sin \Delta_i), \quad (3b)$$

23

1 where dX_{ir} and dY_{ir} are reduced differential components of
 2 the i 'th received vector. Similarly, the quadrature
 3 components of the received signal X_i and Y_i may be directly
 4 reduced to the corresponding components of the reference
 5 vector:

$$7 \quad X_{ir} = (A_0/a_i)(X_i \cos \Delta_i - Y_i \sin \Delta_i), \quad (3c)$$

$$8 \quad Y_{ir} = (A_0/a_i)(Y_i \cos \Delta_i + X_i \sin \Delta_i). \quad (3d)$$

9
 10 Thus, the reduction procedure can be described by equations
 11 3(a)–3(d) or by a corresponding table. For example, a QPSK
 12 system may have the constellation vectors $X_{c1} = -1, Y_{c1} = -1$;
 13 $X_{c2} = -1, Y_{c2} = 1$; $X_{c3} = 1, Y_{c3} = -1$; $X_{c4} = 1, Y_{c4} = 1$, which are
 14 typical for many wireless applications. For this example,
 15 one of the constellation vectors should be assigned as the
 16 reference vector, because in this case, the phase
 17 difference Δ_i between the decision vector and the reference
 18 vector is a multiple of $\pi/2$. If, for example, the
 19 reference vector is $X_0 = 1, Y_0 = 1$, the reduction procedure may
 20 be described by Table 1:

21

22

23

1 Table 1.

Decision Vector	Δ_i	dX_{ir}	dY_{ir}	X_{ir}	Y_{ir}
(-1,1)	$3\pi/2$	dY_i	$-dX_i$	Y_i	$-X_i$
(-1,-1)	π	$-dX_i$	$-dY_i$	$-X_i$	$-Y_i$
(1,-1)	$\pi/2$	$-dY_i$	dX_i	$-Y_i$	X_i
(1,1)	0	dX_i	dY_i	X_i	Y_i

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As one can see, the reduction procedure in this particular case does not need any calculations. In a similar manner, more complicated tables for reduction of the received signals in multiposition QAM systems may be generated.

It should be appreciated by those skilled in the art, that the reduced coordinates of equations (3a)-(3d) may be meaningfully averaged (in contrast to coordinates dX_i and dY_i which would typically average to zero). According to the invention, the reduced coordinates are averaged for a

1 given sequence of N symbols, defined by indexes from (i-N)
 2 to i, as follows:

3

$$4 \quad dX_r(i) = (1/N) \sum dX_{jr} = (A_0/N) * \sum_{j=i-N}^i (dX_j \cos \Delta_j - dY_j \sin \Delta_j) / a_j, (4a)$$

$$5 \quad dY_r(i) = (1/N) \sum dY_{jr} = (A_0/N) * \sum_{j=i-N}^i (dY_j \cos \Delta_j + dX_j \sin \Delta_j) / a_j, (4b)$$

6

7 where $dX_r(i)$ and $dY_r(i)$ are averaged differential components
 8 at the i'th received symbol. Values $dX_r(i)$ and $dY_r(i)$ from
 9 equations (4a) and (4b) are the current estimates of
 10 coordinates of differences between the reference vector and
 11 the shifted reference vector in the (x,y) space. They are
 12 the basis for carrier phase tracking.

13

14 Similarly, the reduced quadrature components of the
 15 received signal X_{ir} and Y_{ir} from equations (3c) and (3d) may
 16 be averaged:

17

$$18 \quad X_r(i) = (1/N) \sum X_{jr} = (A_0/N) * \sum_{j=i-N}^i (X_j \cos \Delta_j - Y_j \sin \Delta_j) / a_j, (4c)$$

$$19 \quad Y_r(i) = (1/N) \sum Y_{jr} = (A_0/N) * \sum_{j=i-N}^i (Y_j \cos \Delta_j + X_j \sin \Delta_j) / a_j, (4d)$$

20

1 Values $X_r(i)$ and $Y_r(i)$ from equations (4c) and (4d) are the
2 current estimates of coordinates of the shifted reference
3 vector in the (x,y) space. They can be also used as the
4 basis for carrier phase tracking.

5

6 It should be appreciated by those skilled in the art
7 that the averaging of equations (4a)-(4d) can be
8 implemented in different manners. One manner of
9 implementation is the conventional averaging with a sliding
10 window. In this case, the estimates $dX_r(i)$ and $dY_r(i)$, as
11 well as $X_r(i)$ and $Y_r(i)$, are calculated for each symbol by
12 averaging the N preceding symbols. This approach
13 guarantees the most accurate phase correction, but it
14 requires considerable processing resource and memory. This
15 level of phase correction may not always be deemed
16 necessary in typical wireless systems with slow phase
17 changes.

18

19 A second manner of implementing equations (4a)-(4d)
20 is to average blocks of N symbols. In this case the
21 estimates dX_r and dY_r , as well as X_r and Y_r , are calculated
22 for each block of N symbols (block by block), and phase
23 correction is provided once per N -symbol block. This

1 approach needs very little memory and requires minimal
2 processing.

3

4 It should be also noted that if all decisions
5 participating in any of the averaging procedures of
6 equations (4a)-(4d) are correct, then the generated
7 estimate is an optimal one, i.e., it is unbiased and
8 effective in terms of the minimum variance. In other
9 words, averaging reduced signal components and averaging
10 reduced differences between the received signals and
11 decisions provide equivalent nonbiased and efficient
12 estimates of the phase shift.

13

14 This statement is illustrated in Fig. 2, which shows
15 results of a simulation of equations (3) and (4). The
16 constellation points are indicated by small crosses and the
17 reference vector in this example has coordinates (2,0).
18 Randomly transmitted constellation points are phase shifted
19 by $\pi/11$ and the Gaussian noise is added. Received signals
20 are indicated by stars and combined into clusters of
21 received vectors. Then all received vectors and
22 differential vectors are reduced (via rotation) and
23 transformed into a cluster of reduced received vectors and
24 into a cluster of reduced received differential vectors.

1 Both transformations were carried out for an error-free
2 decision. It will be appreciated that the two resulting
3 clusters are congruent and may be one-to-one converted from
4 one to another by shifting their X-coordinates by 2.
5 Results of the averaging of these clusters are indicated by
6 circles with points at their centers; naturally, they
7 differ exactly by 2 in the X-coordinate.

8

9 As previously mentioned, erroneous decisions cause
10 bias in the estimate, and for a large error rate this bias
11 may be considerable. Fig. 3 and Fig. 4 qualitatively
12 illustrate this effect.

13

14 Fig. 3 and Fig. 4 show simulation results for
15 equations (3) and (4) at a phase shift $\pi/16$ and a SNR
16 corresponding to (the relatively large) symbol error rate
17 SER=0.01. Fig. 3, however, corresponds to error-free
18 decisions, while Fig. 4 corresponds to the decisions which
19 include errors with the above SER. By comparing Figs. 3
20 and 4, it is seen that clusters of reduced signals which
21 include erroneous decisions are more dispersed than the
22 clusters of reduced signals which involve error-free
23 decisions.

24

1 The algorithms of equations (3) and (4) have been
2 simulated and computer tested to estimate their efficiency
3 for error-free and erroneous decisions. In addition, the
4 method of the invention of phase estimation based on
5 averaging coordinates of the reduced differential received
6 signal, was compared with the existing prior art method
7 which is based on averaging phase shift of the received
8 signals. In the test, the simulation program shifted by
9 $\pi/20$ the phases of transmitted 16-QAM random signals. Both
10 compared methods were simulated in parallel with a 100-
11 symbol averaging interval and with different SNRs,
12 corresponding to symbol error rates (SER) 0.01, 0.05 and
13 0.1.

14
15 The result of the test can be briefly described as
16 follows. For error-free decisions both methods provide
17 unbiased estimates of the phase shift, and dispersion of
18 the phase estimates increases with increasing SER. For the
19 100-symbol averaging interval, the mean deviation lies
20 within the limits of 0.6° - 1.2° , depending on the SER.
21 However, the method of the invention provides less
22 dispersion of phase estimates. In particular, the method
23 of the invention gains 2% in the phase estimate dispersion
24 at SER=0.01, 5% at SER=0.05, and 10% at SER=0.1.

1
2 For decisions that include errors, both methods
3 provide approximately the same bias in the phase estimates,
4 and the bias increases with increasing SER. For the
5 considered conditions and phase shift $\pi/20=9^\circ$, real phase
6 shift estimates were equal to 8.5° at $SER=0.01$ (-0.5° bias
7 or 6%), 7.5° at $SER=0.05$ (-1.5° bias or 17%), and 5.5° at
8 $SER=0.1$ (-3.5° bias or 39%). In addition, dispersion of the
9 phase estimates increases with increasing SER. For the
10 100-symbol averaging interval, the mean deviation lies
11 within the same limits 0.6° - 1.2° , depending on SER. The
12 method of the invention provides the minimum dispersion of
13 phase estimates. Compared with the prior art method, the
14 method of the invention gains 1.5% in the phase estimate
15 dispersion at $SER=0.01$, 2.5% at $SER=0.05$, and 3.5% at
16 $SER=0.1$.

17
18 It is clear from the simulation that at severe channel
19 conditions ($SER>0.01$), it is desirable to correct
20 estimates. A simple method of estimate correction is to
21 exclude extreme points in the cluster of reduced signals;
22 and these extreme points can be easily identified because,
23 as one can see from comparing Fig. 3 and Fig. 4,
24 utilization of erroneous decision moves the points to the

1 edge of the cluster. Unfortunately, exclusion of those
2 extreme points can be implemented only after completion of
3 the cluster calculation; i.e., it delays final estimation.

4

5 According to one aspect of the invention, two
6 practical approaches to solving this matter include
7 correction of the final estimate, and exclusion of
8 unreliable points. First, with respect to correction of
9 the final estimate, as it was shown in the stochastic
10 simulation, the estimate bias is a function of three
11 parameters: SER, constellation size, and the mean of the
12 estimate. All three parameters, as a rule, are known
13 during the estimation procedure. For example, for 16-QAM
14 encoding, a phase shift estimate should be increased by 6%
15 at SER=0.01, 17% at SER=0.05, and 39% at SER=0.1. The
16 required function can be determined in advance by means of
17 stochastic simulation of the system for different expected
18 conditions. Unfortunately, the method guarantees good
19 results only for comparatively small phase shifts.

20

21 Exclusion of unreliable points is a more general
22 approach, and it does not need preliminary simulation. The
23 essence of this method is the exclusion of unreliable
24 symbols from the averaging process. As will be appreciated

1 by those skilled in the art, calculation of the reliability
2 of the received symbols is one of functions of soft
3 decision decoder. The corresponding procedure, based on
4 differential components of the received signal, was
5 disclosed in the previously incorporated patent
6 applications. The estimates of symbol reliability can be
7 used for exclusion of symbols, which likely cause the phase
8 estimate bias. In practice, the procedure for excluding
9 unreliable points would include comparing the symbol
10 reliability calculated in the soft decoder with some
11 predetermined threshold.

12

13 In any case, estimates such as set forth above in
14 equations (4) with the proper correction can be used for
15 current decision-making. According to different
16 embodiments of the invention, these estimates can be
17 utilized to correct the received signal, or can be utilized
18 to correct the constellation points.

19

20 Correction of the received signal according to a first
21 embodiment of the invention is typically preferable from an
22 implementation point of view. Correction of the received
23 signal involves the proper rotation of the received signal
24 (correction of the received coordinates) with further

1 decision making, based on the corrected received signal
 2 without changing constellation points.

3

4 In particular, let X_{ic} and Y_{ic} be corrected coordinates
 5 of the received signal (X_i, Y_i) . The coordinates may be
 6 calculated as follows:

7

$$8 \quad X_{ic} = (X_i \cos \phi - Y_i \sin \phi), \quad (5a)$$

$$9 \quad Y_{ic} = (Y_i \cos \phi + X_i \sin \phi), \quad (5b)$$

10

11 where ϕ is a carrier phase shift, which in turn is equal to
 12 a current estimate of a phase difference between the
 13 initial reference vector and corrected (estimated)
 14 reference vector.

15

16 Taking into account that the corrected reference
 17 vector has coordinates $X_0 + dX_r$ and $Y_0 + dY_r$, where dX_r and dY_r
 18 are averaged differential components according to (4), the
 19 trigonometric functions of phase ϕ are derived as follows:

20

$$21 \quad A \sin \phi = (X_0 + dX_r)Y_0 - (Y_0 + dY_r)X_0 = dX_r Y_0 - dY_r X_0, \quad (6a)$$

$$22 \quad A \cos \phi = (X_0 + dX_r)X_0 + (Y_0 + dY_r)Y_0 = (A_0)^2 + dX_r X_0 + dY_r Y_0, \quad (6b)$$

23

24 where

1

$$2 \quad A = A_0((X_0+dX_r)^2 + (Y_0+dY_r)^2)^{0.5}. \quad (6c)$$

3

4 Thus, for example, if the reference signal has coordinates
 5 $X_0=1$ and $Y_0=0$, then $A\sin\phi = -dY_r$ and $A\cos\phi = 1 + dX_r$.

6

7 By substituting equations (6) into (5), the following
 8 expressions are obtained for corrected coordinates of the
 9 received signal:

10

$$11 \quad X_{ic} = (1/A)\{((A_0)^2 + dX_rX_0 + dY_rY_0)X_i - (dX_rY_0 - dY_rX_0)Y_i\}, \quad (7a)$$

$$12 \quad Y_{ic} = (1/A)\{((A_0)^2 + dX_rX_0 + dY_rY_0)Y_i + (dX_rY_0 - dY_rX_0)X_i\}, \quad (7b)$$

13

14 where dX_r and dY_r are the estimates (4) of differential
 15 components of the reference signal.

16

17 In the same manner, corrected coordinates of the
 18 received signal X_{ic} and Y_{ic} can be derived using estimates of
 19 coordinates (4c,d) of the shifted reference vector as
 20 follows:

21

$$22 \quad X_{ic} = (1/A)(X_i(X_rX_0 + Y_rY_0) - Y_i(X_rY_0 - Y_rX_0)), \quad (7c)$$

$$23 \quad Y_{ic} = (1/A)(Y_i(X_rX_0 + Y_rY_0) + X_i(X_rY_0 - Y_rX_0)), \quad (7d)$$

24

1 where $A = A_0((X_r)^2 + Y_r^2)^{0.5}$.

2

3 Equations (7a) - (7d) can be simplified by the proper
4 choice of the reference signals. If, for example, the
5 reference signal has coordinates $X_0=1$ and $Y_0=0$, and account
6 is taken that in this case $A \approx 1$, the following simple
7 expressions are derived from equations (7a) and (7b):

8

$$9 \quad X_{ic} = X_i + (dX_r X_i + dY_r Y_i), \quad (8a)$$

$$10 \quad Y_{ic} = Y_i + (dX_r Y_i - dY_r X_i). \quad (8b)$$

11

12 In this case, correction of the received signal comprises
13 adding of the convolutions in the parentheses to the
14 received components X_i and Y_i ; and thus, in practice,
15 implementation of equations (8a) and (8b) is preferable.

16

17 For the same conditions equations (7c) and (7d) are
18 transformed as follows:

19

$$20 \quad X_{ic} = X_i X_r + Y_i Y_r, \quad (8c)$$

$$21 \quad Y_{ic} = Y_i X_r - X_i Y_r. \quad (8d)$$

22

23 Given all of the above, according to the first
24 embodiment of the invention, the method for correcting the

1 received signal is as follows (the method being described
2 in parallel for both the preferred first embodiment
3 utilizing differential quadrature components and the
4 alternative first embodiment utilizing the quadrature
5 components of the received signal):

6

7 a) the received signal (X_i, Y_i) is corrected with
8 estimates of the differential reference vector (dX_r, dY_r) or
9 with estimates of reference vector (X_r, Y_r) according to
10 equations (7) or (8);

11 b) the corrected received signal (X_{ic}, Y_{ic}) is used for
12 making a decision and for calculating the differential
13 quadrature components of the corrected received signal dX_i
14 and dY_i ;

15 c) using the decision, the differential components dX_i
16 and dY_i or corrected components X_{ic} and Y_{ic} are transformed
17 into the reduced differential components dX_{ir} and dY_{ir} or
18 reduced components X_{ir} and Y_{ir} according to equations (3);

19 d) sequences of reduced differential components dX_{ir}
20 and dY_{ir} or reduced components X_{ir} and Y_{ir} are averaged
21 according to equations (4) to provide a current estimate of
22 the differential reference vector (dX_r, dY_r) or the
23 reference vector (X_r, Y_r) ; and

1 e) upgraded coordinates of the differential reference
2 vector dX_r and dY_r or the reference vector X_r and Y_r are used
3 for the next correction of the received signal according to
4 equations (7) or (8).

5
6 Turning now to Fig. 5, a flow chart of the first
7 embodiment is provided that illustrates the five-step
8 procedure of signal demapping and received signal
9 correction based on estimates of differential components of
10 the reference signal. In the flow chart, certain blocks
11 are shown with a bold outline while other blocks are shown
12 with a thin outline. The blocks shown in the thin outline
13 (e.g., the current decision unit 104, differences
14 calculation unit 108, hard decoder 106, soft decoder 112
15 and channel estimation unit 110) are blocks which are
16 conventional parts of a receiver and are a universal tool
17 of optimal signal processing, including channel estimation
18 and soft decoding, whereas the blocks shown in bold are
19 added for implementing the invention.

20

21 According to the first embodiment of the invention,
22 and as seen in Fig. 5, the received signal is first
23 corrected at 102 using the current estimates of
24 differential reference vector. Then the corrected received

1 signal is utilized for making a decision at 104 (see
2 equation (1)), and the decision is fed to the hard decoder
3 106 and to the differences calculation unit 108. The
4 differences (see equation (2)) are provided for channel
5 estimation at 110 and the soft decoder 112. The current
6 decision 104 determines parameters of signal reduction Δ_i ,
7 A_0, a_i or one or more indications thereof such as A_0/a_i which
8 are stored in the parameters memory 114. Based on these
9 parameters, the differential components of the received
10 signals as determined by the differences calculation unit
11 108 are reduced at 116 (according to equations (3)) and
12 then averaged at 118 (according to equations (4)).
13 Exclusion of unreliable symbols (if applied) is carried out
14 at 120 and is used to eliminate unreliable symbols from the
15 differential signal reduction block prior to their use in
16 the signal averaging block 118. The symbol exclusion block
17 120 utilizes information regarding symbol reliability from
18 the soft decoder 112. Finally, the estimates of
19 coordinates of the differential reference vector as
20 determined by the signal averaging block 118 are fed to the
21 received signal correction block 102.

22

23 It should be noted that the system and method
24 implemented in Fig. 5 does not use any nonlinear

1 calculation of the carrier phase, and all calculation
 2 procedures in the loop of Fig. 5 include only linear
 3 operations, based on carrier projections. An advantage of
 4 the system and method of Fig. 5 is that it does not need
 5 any correction of the constellation points, and, as a
 6 result, preserves the simplest decision making procedure,
 7 based on comparison of the received coordinates with a
 8 limited number of thresholds.

9

10 A second embodiment of utilization of phase shift
 11 estimates is the proper rotation of the constellation
 12 points (correction of the constellation point coordinates)
 13 with further decision making, based on the corrected
 14 constellation points.

15

16 More particularly, let $X_{cn}(i)$ and $Y_{cn}(i)$ be current
 17 corrected coordinates of the constellation points, where
 18 $n=1,2, \dots, m$, and where m represents the number of the
 19 constellation points. With X_0 and Y_0 being coordinates of
 20 the current reference point, the corrected coordinates may
 21 be calculated as follows:

22

$$23 \quad X_{cn}(i) = (A_n/A_0)\{(X_0 + dX_r(i))\cos\theta_n - (Y_0 + dY_r(i))\sin\theta_n\}, \quad (9a)$$

$$24 \quad Y_{cn}(i) = (A_n/A_0)\{(Y_0 + dY_r(i))\cos\theta_n + (X_0 + dX_r(i))\sin\theta_n\}, \quad (9b)$$

1 where θ_n is the phase difference between the reference
 2 vector and the n'th constellation point.

3

4 Equations (9) describe one step of correction of the
 5 constellation points coordinates. During the adaptation
 6 process, $(X_0 \cos \theta_n - Y_0 \sin \theta_n)$ and $(Y_0 \cos \theta_n + X_0 \sin \theta_n)$ can be
 7 considered as estimates of constellation points at the
 8 previous step; i.e.,

9

$$10 \quad X_{cn}(i-1) = (A_n/A_0)(X_0 \cos \theta_n - Y_0 \sin \theta_n), \quad (10a)$$

$$11 \quad Y_{cn}(i-1) = (A_n/A_0)(Y_0 \cos \theta_n + X_0 \sin \theta_n). \quad (10b)$$

12

13 Combining equations (9) and (10) yields:

14

$$15 \quad X_{cn}(i) = X_{cn}(i-1) + (A_n/A_0)(dX_r(i) \cos \theta_n - dY_r(i) \sin \theta_n), \quad (11a)$$

$$16 \quad Y_{cn}(i) = Y_{cn}(i-1) + (A_n/A_0)(dY_r(i) \cos \theta_n + dX_r(i) \sin \theta_n). \quad (11b)$$

17

18 Similarly, corrected coordinates of the constellation
 19 points X_{cn} and Y_{cn} can be derived using coordinates from
 20 equations (4c) and (4d) of the shifted reference vector as
 21 follows:

22

$$23 \quad X_{cn}(i) = (A_n/A_0)(X_r(i) \cos \theta_n - Y_r(i) \sin \theta_n), \quad (11c)$$

$$24 \quad Y_{cn}(i) = (A_n/A_0)(Y_r(i) \cos \theta_n + X_r(i) \sin \theta_n). \quad (11d)$$

1

2 Equations (11a) - (11d) can be significantly
 3 simplified for BPSK and QPSK systems. If, for example, in
 4 a QPSK system with constellation vectors $X_{c1}=-1, Y_{c1}=-1$;
 5 $X_{c2}=-1, Y_{c2}=1$; $X_{c3}=1, Y_{c3}=-1$; $X_{c4}=1, Y_{c4}=1$, the reference
 6 vector is $X_0=1, Y_0=1$, then equations (11a) and (11b) have
 7 the following simple expressions:

8

9 for $n=1,2,3$

10

$$11 \quad X_{cn}(i) = X_{cn}(i-1) \pm dY_r(i), \quad (12a)$$

$$12 \quad Y_{cn}(i) = Y_{cn}(i-1) \pm dX_r(i); \quad (12b)$$

13

14 for $n=4$

15

$$16 \quad X_{cn}(i) = X_{cn}(i-1) + dX_r(i), \quad (12c)$$

$$17 \quad Y_{cn}(i) = Y_{cn}(i-1) + dY_r(i). \quad (12d)$$

18

19 Given all of the above, according to a second
 20 embodiment of the invention, the method for the correction
 21 of constellation points is as follows (the algorithm is
 22 described in parallel for both the second embodiment
 23 utilizing differential quadrature components and an

1 alternative second embodiment utilizing quadrature
2 components of the received signal):

3

4 a) the received signal (X_i, Y_i) is used for making
5 decision, and differential quadrature components of the
6 received signal dX_i and dY_i are calculated according to
7 equations (2);

8 b) using the decision, the differential components dX_i
9 and dY_i or components X_i and Y_i are transformed into the
10 reduced differential components dX_{ir} and dY_{ir} according to
11 equations (3a) and (3b) or into reduced components X_{ir} and
12 Y_{ir} according to equations (3c) and (3d);

13 c) sequences of reduced differential components dX_{ir}
14 and dY_{ir} or reduced components X_{ir} and Y_{ir} are averaged to
15 provide current estimates of the differential reference
16 vector (dX_r, dY_r) according to equations (4a) and (4b) or
17 the reference vector (X_r, Y_r) according to equations (4c)
18 and (4d);

19 d) based on estimates dX_r and dY_r or estimates X_r and
20 Y_r , corrected coordinates of the constellation points X_{cn} and
21 Y_{cn} are calculated according to equations (11); and

22 e) upgraded coordinates of the constellation points X_{cn}
23 and Y_{cn} are used for making the next decision.

24

1 Turning now to Fig. 6, a flow chart is provided that
2 illustrates the above five-step procedure of signal
3 demapping and the constellation points correction according
4 to the second embodiment of the invention, based on
5 estimates of the differential reference components. In the
6 block-diagram, certain blocks are shown in bold lines and
7 certain blocks are shown in thin lines. The blocks shown
8 in the thin outline (e.g., the current decision unit 204,
9 differences calculation unit 208, hard decoder 206, soft
10 decoder 212 and channel estimation unit 210) are blocks
11 which are conventional parts of a receiver and are a
12 universal tool of optimal signal processing, including
13 channel estimation and soft decoding, whereas the blocks
14 shown in bold are added for implementing the invention.

15

16 As seen in Fig. 6, the received signal, first, is used
17 for making a current decision at 204, and the decision is
18 fed to the hard decoder 206 and to the differences
19 calculation unit 208. The differences calculations are
20 used by channel estimation 210 and soft decoder 212. The
21 current decision is also used to determine parameters of
22 signal reduction such as Δ_i , A_0 , a_i , or indications thereof
23 such as A_0/a_i , which are stored in the parameters memory

1 214. Based on these parameters, the differential
2 components of the received signals are reduced at 216 and
3 then averaged at 218. As was described above with
4 reference to Fig. 5, using information provided by the soft
5 decoder 212, exclusion of unreliable symbols (if applied)
6 is carried out at 220 so that only reliable symbols are
7 provided to the signal averaging block 218. Finally,
8 corrected constellation points are calculated at 225, and
9 the upgraded constellation coordinates are fed to the
10 current decision block 204.

11

12 The advantage of the system of Fig. 6 is that it
13 provides optimal adaptive processing without any changing
14 of the received signal. In other words, in the second
15 embodiment of the invention, the receiver does not spend
16 processing time for transformation of each received symbol,
17 and all processing relates only to constellation point
18 correction. The constellation point correction need not be
19 carried out as frequently as the symbol rate; e.g., it can
20 be carried out one time per 100 symbols. The advantage of
21 the system of Fig. 6 may therefore be considerable for
22 small size constellations; for example, for BPSK or QPSK
23 modulation techniques. However, in the case of large size
24 constellations (16-QAM, 64-QAM and so on) the system in

1 Fig. 6 has the disadvantage of requiring a relatively
2 complicated decision making procedure, which includes
3 comparison of the received signal with all upgraded
4 constellation points and recalculation of large number of
5 constellation points. Thus, for the multiposition QAM
6 modulation, the embodiment of Fig. 5 which utilizes
7 correction of the received signal is presently the more
8 preferred method.

9

10 The previously disclosed embodiments provide a general
11 method of phase shift compensation in single carrier and
12 multicarrier pilotless wireless systems with uncorrelated
13 between-carriers phase shifts. In the multicarrier case,
14 the algorithms of the embodiments can provide individual
15 phase tracking for each carrier. However, the algorithms
16 may be simplified for multicarrier wireless system with
17 correlated carriers.

18

19 Completely correlated carriers are found in wireless
20 systems with small carrier diversity and/or with short
21 communication sessions (short packet transmission). Such
22 conditions allow the phase adjustment algorithms to be
23 simplified. According to one aspect of the invention, the
24 simplification may be based on substituting averaging in

1 the time domain by averaging in the frequency domain.
2 According to another aspect of the invention, the
3 simplification may be based on utilization of the same
4 phase shift estimate for all carriers.

5
6 The equations applicable to the multicarrier systems
7 and methods with correlated carriers use the same variables
8 as do the previously described embodiments. In order to
9 distinguish the averaging in time and frequency domains,
10 the index "k", which are carrier numbers, will be used
11 instead of the index "i", which were symbol numbers in time
12 domain.

13
14 With that change in designation, the differential
15 components dX_k and dY_k of the k-th carrier, equivalent to
16 differential components in equations (2), are

17
18
$$dX_k = (X_k - X_{dk}) , \quad (13a)$$

19
$$dY_k = (Y_k - Y_{dk}) , \quad (13b)$$

20
21 where X_k and Y_k are the quadrature components of the k-th
22 carrier, and X_{dk} , Y_{dk} are the quadrature components of the
23 k-th carrier decision that typically correspond to the
24 constellation point nearest to the received vector (X_k, Y_k) .

1

2 The reduced differential components dX_{kr} and dY_{kr} of the
3 k-th carrier are

4

$$5 \quad dX_{kr} = (A_0/a_k)(dX_k \cos \Delta_k - dY_k \sin \Delta_k), \quad (14a)$$

$$6 \quad dY_{kr} = (A_0/a_k)(dY_k \cos \Delta_k + dX_k \sin \Delta_k), \quad (14b)$$

7

8 where Δ_k is the phase difference between the decision and
9 reference vectors at the k-th carrier, a_k is the amplitude
10 of the decision vector at the k-th carrier, and A_0 is the
11 amplitude of the reference vector.

12

13 In the same manner that the differential quadrature
14 components are reduced in equations (14a) and (14b), the
15 quadrature components of the received carriers X_k and Y_k may
16 be directly reduced to the corresponding components X_{kr} and
17 Y_{kr} of the reference vector:

18

$$19 \quad X_{kr} = (A_0/a_k)(X_k \cos \Delta_k - Y_k \sin \Delta_k), \quad (14c)$$

$$20 \quad Y_{kr} = (A_0/a_k)(Y_k \cos \Delta_k + X_k \sin \Delta_k). \quad (14d)$$

21

22 From equations (14a) and (14b), it will be appreciated that
23 the averaged reduced differential components dX_r and dY_r are

1

$$2 \quad dX_r = (1/K) \sum dX_{kr} = (A_0/K) * \sum_{k=1}^K (dX_k \cos \Delta_k - dY_k \sin \Delta_k) / a_k, \quad (15a)$$

$$3 \quad dY_r = (1/K) \sum dY_{kr} = (A_0/K) * \sum_{k=1}^K (dY_k \cos \Delta_k + dX_k \sin \Delta_k) / a_k, \quad (15b)$$

4

5 where K is the number of carriers. Similarly, the reduced
6 quadrature components of the received carriers X_{kr} and Y_{kr} as
7 set forth in equations (14c) and (14d) may be averaged:

8

$$9 \quad X_r = (1/K) \sum X_{kr} = (A_0/K) * \sum_{k=1}^K (X_k \cos \Delta_k - Y_k \sin \Delta_k) / a_k, \quad (15c)$$

$$10 \quad Y_r = (1/K) \sum Y_{kr} = (A_0/K) * \sum_{k=1}^K (Y_k \cos \Delta_k + X_k \sin \Delta_k) / a_k. \quad (15d)$$

11

12 The estimates of the corrected differential reference
13 signal (equations (15a) and (15b)) or the corrected
14 reference signal (equations (15c) and (15d)) may be
15 utilized for correction of a common carrier phase shift in
16 the same manner as described above with reference to
17 estimate equations (4a) - (4d). However, it should be
18 noted that, in contrast to estimates of equations (4) which
19 provide an individual estimate for each carrier, the
20 estimates provided by equations (15) are the same for all
21 carriers. Therefore, estimate (15) can be used for

1 correction of all received carriers or for correction of
 2 constellation points for all carriers.

3

4 In correcting the received carriers, the procedure is
 5 generally equivalent to equations (7), and can be described
 6 as follows for differential quadrature components of
 7 carriers:

8

$$9 \quad X_{kc} = (1/A) \{ ((A_0)^2 + dX_r X_0 + dY_r Y_0) X_k - (dX_r Y_0 - dY_r X_0) Y_k \}, \quad (16a)$$

$$10 \quad Y_{kc} = (1/A) \{ ((A_0)^2 + dX_r X_0 + dY_r Y_0) Y_k + (dX_r Y_0 - dY_r X_0) X_k \}, \quad (16b)$$

11

12 and as follows for quadrature components of carriers:

13

$$14 \quad X_{kc} = (1/A) (X_k (X_r X_0 + Y_r Y_0) - Y_k (X_r Y_0 - Y_r X_0)), \quad (16c)$$

$$15 \quad Y_{kc} = (1/A) (Y_k (X_r X_0 + Y_r Y_0) + X_k (X_r Y_0 - Y_r X_0)), \quad (16d)$$

16

17 where X_{kc} , Y_{kc} are the corrected quadrature components of the
 18 k-th carrier, X_k , Y_k are the received quadrature components
 19 of the k-th carrier, dX_r and dY_r are the estimates of
 20 differential components of the reference signal calculated
 21 according to equations (15a) and (15b), and X_r and Y_r are
 22 the estimate of components of the reference signal
 23 calculated according to equations (15c) and (15d). Those

1 skilled in the art will appreciate that the expressions in
2 equations (16a) - (16d) are the same for all carriers.

3

4 Given the above, the method of carrier correction in
5 multicarrier systems having correlated phase shifts may be
6 described as follows:

7

8 a) a set of received carriers (X_k, Y_k) is transformed
9 into a set of corrected carriers (X_{kc}, Y_{kc}) using common
10 estimates of differential quadrature components of the
11 reference signal dX_r and dY_r according to equations (16a)
12 and (16b) or quadrature components of the reference signal
13 X_r and Y_r using equations (16c) and (16d) for all carriers;

14 b) the set of corrected carriers (X_{kc}, Y_{kc}) is used for
15 making multicarrier current decisions, and differential
16 quadrature components of the carriers dX_k and dY_k are
17 calculated according to equations (13);

18 c) using the decisions, the set of differential
19 components dX_k and dY_k or the set of components X_k and Y_k are
20 transformed into a set of reduced differential components
21 dX_{kr} and dY_{kr} according to equations (14a) and (14b) or into
22 a set of reduced components X_{kr} and Y_{kr} according to
23 equations (14c) and (14d);

1 d) the set of reduced differential components dX_{kr} and
2 dY_{kr} are averaged according to equations (15a) and (15b), or
3 the set of reduced components X_{kr} and Y_{kr} are averaged
4 according to equations (15c) and (15d) to provide the
5 current estimate of the differential reference vector (dX_r ,
6 dY_r) or reference vector (X_r , Y_r); and

7 e) upgraded coordinates of the differential reference
8 vector dX_r and dY_r or the reference vector X_r and Y_r , common
9 for all carriers, are used for correction of the next
10 multicarrier symbol according to equations (16).

11

12 Turning now to Fig. 7, a flow chart is provided which
13 illustrates the above-described five step procedure of
14 signal demapping and received carriers correction for
15 multicarrier systems with correlated carrier phase shifts,
16 based on differential quadrature components of the
17 carriers. As with Fig. 5, certain blocks are shown with a
18 bold outline while other blocks are shown with a thin
19 outline; with the blocks shown in the thin outline
20 indicating conventional parts of a receiver. Blocks 302 -
21 318 of Fig. 7 are similar to blocks 102 - 118 of Fig. 5
22 (with numbering differing by 200). The difference between
23 the elements of Fig. 5 and Fig. 7 can be explained as
24 follows: the system of Fig. 5 which utilizes equations (3),

1 (4), (7) and (8) provides an individual phase shift
 2 estimate for each carrier on the basis of averaging each
 3 carrier's signals in the time domain, whereas the system of
 4 Fig. 7 which utilizes equations (13) - (16) provides a
 5 common phase shift estimate for all carriers on the basis
 6 of averaging carrier signals in the frequency domain.

7
 8 It should be appreciated by those skilled in the art
 9 that the second embodiment of the invention which is
 10 directed to correcting constellation point coordinates can
 11 be used in conjunction with the discussion above regarding
 12 multicarrier systems having correlated phase shifts. In
 13 particular, in the case of constellation points correction
 14 the procedure is equivalent to equations (11), and can be
 15 described as follows for differential components dX_r , dY_r of
 16 the reference vector:

17

$$18 \quad X_{cn} = X_n + (A_n/A_0)(dX_r \cos \theta_n - dY_r \sin \theta_n); \quad (17a)$$

$$19 \quad Y_{cn} = Y_n + (A_n/A_0)(dY_r \cos \theta_n + dX_r \sin \theta_n); \quad (17b)$$

20

21 and as follows for components X_r , Y_r of the reference
 22 vector:

23

$$24 \quad X_{cn} = (A_n/A_0)(X_r \cos \theta_n - Y_r \sin \theta_n), \quad (17c)$$

$$Y_{cn} = (A_n/A_0)(Y_r \cos \theta_n + X_r \sin \theta_n), \quad (17d)$$

2

3 where (X_{cn}, Y_{cn}) is the corrected n-th constellation point,
 4 and (X_n, Y_n) is the initial n-th constellation point. Again,
 5 it should be emphasized that the corrected constellation
 6 point (X_{cn}, Y_{cn}) in equations (17) is the same for all
 7 carriers.

8

9 Given the above, the complete algorithm of
 10 constellation point correction in multicarrier systems may
 11 be described as follows:

12

13 a) a set of received carriers (X_k, Y_k) is used for
 14 making multicarrier current decisions, and a set of
 15 differential quadrature components of the received carriers
 16 dX_k and dY_k are calculated according to equations (13);

17 b) using the decisions, the set of differential
 18 components dX_k and dY_k or the set of components X_k and Y_k are
 19 transformed into a set of reduced differential components
 20 dX_{kr} and dY_{kr} according to equations (14a) and (14b) or into
 21 a set of reduced components X_{kr} and Y_{kr} according to
 22 equations (14c) and (14d);

23 c) the set of reduced differential components dX_{kr} and
 24 dY_{kr} or the set of reduced components X_{kr} and Y_{kr} are averaged

1 to provide current estimates of the differential reference
2 vector (dX_r , dY_r) according to equations (15a) and (15b) or
3 of the reference vector (X_r , Y_r) according to equations
4 (15c) and (15d);

5 d) based on estimates dX_r and dY_r or estimates X_r and
6 Y_r , corrected coordinates of the constellation points X_{cn} and
7 Y_{cn} are calculated according to equations (17); and

8 e) upgraded coordinates of the constellation points X_{cn}
9 and Y_{cn} , which are the same for all carriers, are used for
10 making the next multicarrier decision.

11

12 Turning now to Fig. 8, a flow chart is provided which
13 illustrates the above-described five step procedure of
14 signal demapping and constellation points correction for
15 multicarrier systems with correlated carrier phase shift,
16 based on differential quadrature components of the
17 carriers. It will be appreciated that the flow chart of
18 Fig. 8 includes blocks 404 - 425 which are similar to
19 blocks 204 - 425 described above with reference to Fig. 6.
20 The difference between the two is that the system of Fig. 6
21 provides individual correction of the constellation for
22 each carrier on the basis of averaging each carrier signals
23 in the time domain, whereas the system of Fig. 8 provides a

1 common constellation for all carriers on the basis of
2 averaging carrier signals in the frequency domain.

3

4 It should be noted that in the case of correlated-
5 carrier phase shifts, the disadvantage of the constellation
6 point correction as opposed to received signal correction
7 (i.e., the necessity of recalculating a large number of
8 constellation points) is transformed into an advantage.
9 More particularly, in the correlated-carrier phase shift
10 case using signal correction, each carrier must be
11 corrected during each symbol, i.e. the number of
12 corrections per symbol is equal to the number of carriers
13 K . In contrast, in the correlated-carrier phase shift case
14 using constellation point correction, the corrected set of
15 constellation points are common for all carriers, i.e. a
16 number of correction is equal to constellation size " m ".
17 If $m < K$, constellation point correction requires less
18 computation than algorithm signal point correction even
19 when correction is carried out for each symbol. Besides,
20 correction of constellation points can be provided once per
21 $n > 1$ symbols depending on how fast the phase is changing.
22 Therefore, a mean number of corrections per symbol is equal
23 to m/n , which is, as a rule, less than K in wireless
24 systems.

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According to another aspect of the invention, the basic algorithm in the case of the correlated-carrier phase shift may be further modified and simplified. The additional simplification is best understood with reference first to Fig. 9, where a 16-QAM constellation is depicted in (X,Y)-space. Fig. 9 shows two received vectors: received vector 1, and received vector 2. The vectors have the same phase shift relative to the constellation point $(-3,3)$, but with opposite sign: $+f$ and $-f$. Both received vectors provide the same decision; i.e., decision vector $(-3,3)$. Fig. 9 shows differential vectors as differences between the received vectors and the decision vector. If the reference vector is $(1,0)$, resulting reduced received vectors and reduced differential vectors will result as is indicated in Fig. 9, with corrected reference vectors 1 and 2, and reduced differential vectors 1 and 2. By reference to these vectors in Fig. 9, it can be seen that the sign of the Y-coordinates of the reduced differential vectors or corrected reference vectors is the same as the sign of the received vector's phase shift. In addition, the phase shift is proportional to the absolute value of the Y-coordinates of the vectors.

Based on these observations, a general simplified algorithm of phase tracking in a multicarrier system can be mathematically derived. Estimates of the Y-coordinates of the differential reference vector and the reference vector can be presented as follows:

$$dY_r = (A_0 / K) * \sum_{k=1}^K (dY_k \cos \Delta_k + dX_k \sin \Delta_k) / a_k, \quad (18a)$$

$$Y_r = (A_0 / K) * \sum_{k=1}^K (Y_k \cos \Delta_k + X_k \sin \Delta_k) / a_k. \quad (18b)$$

If the reference vector is (1,0), then the estimate (18a) is equal to (18b), and for small phase shift both of them are equal to the shift:

$$\phi \approx dY_r = Y_r. \quad (19)$$

Given the above, a simplified method of carrier correction in multicarrier systems with correlated phase shift may be described as follows:

a) the received carriers are phase corrected by predetermined value dY_r or Y_r radians;

1 b) the set of corrected carriers is used for making
2 multicarrier current decisions, and differential quadrature
3 components of the corrected carriers dX_k and dY_k are
4 calculated according to equations (13);

5 c) using the decisions, the set of differential
6 quadrature components dX_k and dY_k or the set of quadrature
7 components of the carriers X_k and Y_k are reduced and
8 averaged according to equations (18); and

9 d) upgraded estimates (18) are used for the next step
10 of received carriers correction.

11

12 Fig. 10 shows a flow diagram of the simplified
13 carriers correction algorithm. The carrier phase of the
14 received multicarrier signal is first corrected at 602 by
15 dY_r radians. Then the corrected received signals are
16 utilized for making a multicarrier decision at 604
17 according to equations (13), and the decision is fed to the
18 hard decoder 606 and to the differential component
19 calculation unit 608. The differential components are
20 provided for channel estimation at 610 and the soft decoder
21 612. The multicarrier current decision 604 determines
22 parameters of signal reduction a_k and Δ_k which are stored in
23 the parameters memory 614. Based on these parameters, the
24 differential components of the received signals as

1 determined by the differences calculation unit 608 are
 2 reduced at 616 and averaged at 618 (according to equation
 3 (18)) to provide a carrier phase correction signal which is
 4 fed back to the carrier phase correction block 602.

5

6 According to another aspect of the invention, the
 7 phase tracking algorithm for multicarrier systems with
 8 correlated between-carrier phase shifts is further
 9 simplified based on a "majority vote" approach. In this
 10 case the accumulations of terms in (18) are replaced by
 11 accumulation of their signs:

12

$$13 \quad D_{+-} = \sum_{k=1}^K \text{Sign} (dY_k \cos \Delta_k + dX_k \sin \Delta_k) \quad (20a)$$

$$14 \quad D_{+-} = \sum_{k=1}^K \text{Sign} (Y_k \cos \Delta_k + X_k \sin \Delta_k), \quad (20b)$$

15

16 where $\text{Sign}() = +1$ or -1 . The resulting integer D_{+-} is a
 17 difference between the number of carriers with positive
 18 phase shifts and the number of carriers with negative phase
 19 shifts. This integer reflects a carrier "majority vote",
 20 and its sign determines a direction for common phase shift
 21 adjustment.

22

1 It should be noted that replacement of the terms of
2 equations (18) by their signs in equations (20) provides
3 some mitigation of the effect of wrong decisions, because
4 in this case any wrong decision cannot dramatically change
5 the result.

6
7 Additional robustness of the algorithm of equations
8 (20) may be achieved by using a lower bound for majority
9 votes; i.e., if the modulo of D_+ is less than some
10 predetermined threshold T_d , no corrections are provided.
11 Threshold T_d preferably depends on the number of carriers
12 involved. System simulation shows that a threshold equal
13 to 10% of all carriers participating in the adaptation
14 process provides sufficient robustness of the system. For
15 example, $T_d = 5$ for WLAN according to the IEEE 802.11a
16 standard.

17
18 Since integer D_+ from equations (20) determines only a
19 direction of common phase shift adjustment, it will be
20 appreciated that it is also desirable to obtain a
21 quantitative value for the phase shift adjustment.

22

23 According to another aspect of the invention, several
24 methods of determining the phase shift value are provided.

1 A first method comprises averaging projections of the
 2 carrier majority. According to this method, differential
 3 carrier projections or carrier projections are accumulated
 4 as in equations (18), but only for carriers which are from
 5 the majority votes. The resulting value is then divided by
 6 a number of majority carriers. For example, if the total
 7 number of carriers is equal to K , then the number of
 8 majority carriers is equal to $(K + |D_{+}|)/2$. In other words,
 9 in this method the phase shift is corrected by the
 10 projections corresponding to the largest number of
 11 occasions. It should be noted that the method has shown
 12 good results in simulation.

13

14 A second method of determining phase shift value is
 15 based on assumption that the phase shift is small enough
 16 and can be efficiently corrected by changing carrier phases
 17 with a constant small increment. In this case, the phase
 18 adjustment algorithm should determine only a direction of
 19 the adjustment. In turn, the adjustment direction $\text{Sign}(f)$
 20 can be found as a sign of value D_{+} from equations (20):

21

$$22 \quad \text{Sign}(\phi) = \text{Sign} \left(\sum_{k=1}^K \text{Sign} (dY_k \cos \Delta_k + dX_k \sin \Delta_k) \right) \quad (21a)$$

$$\text{Sign}(\phi) = \text{Sign} \left(\sum_{k=1}^K \text{Sign} (Y_k \cos \Delta_k + X_k \sin \Delta_k) \right). \quad (21b)$$

2

3 It should be noted that the method of changing carrier
 4 phases with a constant small increment is a simple one
 5 because it does not require phase shift calculation or a
 6 calculation of the number of majority votes. Its
 7 disadvantage, however, is that it is not as accurate in
 8 providing the constant increment over a wide range of phase
 9 shift changing.

10

11 Generally, the majority algorithm of phase tracking
 12 with constant increment may be described as follows:

13

14 a) all received carriers are phase corrected with some
 15 predetermined phase shift or with constant phase increment
 16 and with some predetermined sign;

17 b) the set of corrected carriers is used for making
 18 multicarrier current decisions (X_{dk}, Y_{dk}) , and differential
 19 quadrature components of the carriers dX_k and dY_k are
 20 calculated according to equations (13);

21 c) using the decisions, the set of differential
 22 components dX_k and dY_k or the set of corrected components X_k

1 and Y_k are reduced and then transformed into an integer $D_{..}$
2 according to a majority vote algorithm (20);
3 d) if $D_{..}$ is less than some predetermined threshold T_d ,
4 no phase correction is provided; otherwise the direction of
5 phase correction is determined by a sign of $D_{..}$, and the
6 phase shift value is taken equal to either the average
7 phase shift of the majority carriers or the predetermined
8 constant increment; and
9 e) the newly determined sign of phase adjustment and
10 phase shift value are used in the next step of carrier
11 phase correction.

12

13 Fig. 11 shows a flow diagram of the phase adjustment
14 system. At 702 the carrier phase of the received
15 multicarrier signal is first corrected with some
16 predetermined phase shift or with a constant phase
17 increment and with some predetermined sign. Then the
18 corrected received signals are utilized for making a
19 multicarrier decision at 704 and the decision is fed to the
20 hard decoder 706 and to the differential component
21 calculation unit 708. The differential components are
22 provided for channel estimation at 710 and the soft decoder
23 712. The multicarrier current decision 704 determines a
24 parameter of signal reduction Δ_k which is stored in the

1 parameters memory 714. Based on this parameter, the
2 differential components of the received signals as
3 determined by the differences calculation unit 708 are
4 reduced at 716, and at 718a a majority vote algorithm is
5 utilized to provide integer D_{+} according to equation (20).
6 At 718b a determination is made as to whether D_{+} is greater
7 than (or equal to) some predetermined threshold T_d . If it
8 is, the direction of phase correction is determined at 718c
9 by a sign of D_{+} , and the phase shift value is taken equal
10 to either the average phase shift of the majority carriers
11 or the predetermined constant increment. If the integer D_{+}
12 is not greater than the predetermined threshold, than at
13 718d no phase correction is provided. The results of
14 blocks 718c or 718d are fed back to block 702 for use in
15 the next step of carrier phase correction.

16

17 It will be appreciated by those skilled in the art,
18 that the above-described algorithms are based on signal
19 correction in the frequency domain because they provide
20 adjustment of carrier quadrature components, which, in
21 their turn, are results of a FFT. This frequency domain
22 approach, i.e. signal correction after FFT, completely
23 solves carrier phase tracking problem in OFDM systems.
24 However, with respect to frequency offset compensation, the

1 frequency domain approach only partly solves the problem.
2 The fact is that in the OFDM systems the frequency offset
3 causes both carrier phase shifts and violation of carrier
4 orthogonality. Violation of carrier orthogonality, in its
5 turn, causes considerable intercarrier interference. The
6 considered algorithms provide phase shift compensation but
7 they cannot eliminate or mitigate the intercarrier
8 interference. To the extent that the interference power is
9 a monotonical function of the frequency offset, the offset
10 compensation after FFT is efficient only for comparatively
11 small frequency shifts.

12

13 In principle, the intercarrier interference may be
14 compensated for in the frequency domain (after FFT) by
15 means of interference cancellation techniques, based on
16 decision feedback. However, this approach is complex,
17 especially for OFDM systems with a large number of
18 carriers.

19

20 Another approach is frequency offset compensation in
21 the time domain before FFT. The time domain approach is
22 attractive because, first, it allows the system to
23 reestablish carrier orthogonality and avoid intercarrier
24 interference, and, second, it may be simply implemented.

1 A general algorithm of frequency offset compensation
2 in the time domain may be derived from the Discrete Fourier
3 Transform theory: if the n-th complex sample of a signal,
4 frequency shifted by Δf Hz, is S_n , then the n-th sample of
5 the unshifted signal is complex number $S_n \exp(-jn\phi)$, where
6 $\phi = 2\pi\Delta f T$ and T is an FFT interval.

7
8 The phase shift ϕ in this algorithm corresponds to the
9 phase shift estimate provided by the previously described
10 algorithms for multicarrier OFDM systems, based on reducing
11 and averaging differential quadrature components of the
12 received carriers. General expressions for trigonometrical
13 function of phase shift ϕ are provided by equations (6),
14 where differential components dX_r and dY_r are calculated
15 according to equations (15a) and (15b). A simplified
16 algorithm of phase shift estimation can be also utilized to
17 determine the phase shift ϕ for frequency offset
18 compensation in time domain.

19
20 Turning now to yet another aspect of the invention, a
21 per-carrier adaptive equalizer for multicarrier wireless
22 systems is provided, and uses estimates of differential
23 quadrature components of the reference vector.

24

1 As previously mentioned, the proposed method of
2 carrier phase tracking can be utilized for adaptive
3 equalization of received multicarrier signals. Generally,
4 in multicarrier systems the equalizer function includes
5 adjustment of amplitudes and phases of all received
6 carriers to the corresponding reference signals (which are
7 ideally the constellation points). As a rule, wireless
8 systems have a special training signal (preamble), which is
9 used for preliminary equalization of all carriers. At the
10 end of preamble the equalizer is "frozen" and during the
11 data transmission session each received carrier is
12 equalized by means of convolution with some predetermined
13 constant vector. For purposes herein, this preliminary
14 equalizer will be called a "static equalizer", which
15 emphasizes the fact that during data transmission it does
16 not change equalization parameters. However, in channels
17 with variable parameters, amplitudes and phases of the
18 carriers fluctuate during the session, and the static
19 equalizer does not provide perfect correction of the
20 received signals. So in many cases, wireless systems
21 require adaptive equalization during the communication
22 session to provide perfect coherent signal processing. For
23 purposes herein, the equalizer which implements the
24 adaptive equalization is called a "dynamic equalizer",

1 which emphasizes the fact that during data transmission it
 2 does adjust equalization parameters to the channel
 3 conditions.

4

5 Frequency offset compensation and phase shift tracking
 6 may be considered part of the adaptive equalization
 7 process. The corresponding algorithms, based on estimates
 8 of differential quadrature components of the reference
 9 vector, were considered above. According to this aspect of
 10 the invention, the same approach is taken for realization
 11 of the frequency equalizer function as a whole.

12

13 In particular, let X_k and Y_k be quadrature components
 14 of the k -th carrier at the output of the static equalizer;
 15 i.e., they are a preliminarily equalized received signal,
 16 corresponding to the k -th carrier. Further, assume that
 17 the equalized signal (X_k, Y_k) has changed both its amplitude
 18 and phase compared to the initial equalization during the
 19 preamble. Now, if the k -th carrier phase shift is equal to
 20 ϕ_k , then the phase-corrected coordinates of the k -th
 21 received carrier X_{kc} and Y_{kc} may be calculated as follows:

22

$$23 \quad X_{kc} = X_k \cos \phi_k - Y_k \sin \phi_k, \quad (22a)$$

$$24 \quad Y_{kc} = Y_k \cos \phi_k + X_k \sin \phi_k. \quad (22b)$$

1

2 The coordinate of equations (22) correspond to the proper
3 rotation of the received vector without changing its
4 amplitude.

5

6 Assume now that the relative change of the amplitude
7 is equal to δA_k ; in other words δA_k is a ratio of the initial
8 carrier amplitude to the new carrier amplitude. Then, phase
9 and amplitude corrected (equalized) coordinates of the k-th
10 received carrier X_{ke} and Y_{ke} may be calculated as follows:

11

$$12 \quad X_{ke} = \delta A_k (X_k \cos \phi_k - Y_k \sin \phi_k), \quad (23a)$$

$$13 \quad Y_{ke} = \delta A_k (Y_k \cos \phi_k + X_k \sin \phi_k). \quad (23b)$$

14

15 To provide equalization according to equations (23), values
16 must be determined for δA_k and ϕ_k .

17

18 The carrier phase shift ϕ_k is equal to a current
19 estimate of the phase difference between the reference
20 vector and corrected (estimated) reference vector. Taking
21 into account equations (6), trigonometrical functions of
22 the phase shift ϕ_k can be derived as follows:

23

24

$$\sin\phi_k = (dX_{rk} * Y_0 - dY_{rk} * X_0) / B_k, \quad (24a)$$

$$\cos\phi_k = ((A_0)^2 + dX_{rk} * X_0 + dY_{rk} * Y_0) / B_k, \quad (24b)$$

3

4 where dX_{rk} and dY_{rk} are estimates of the differential
 5 quadrature components of the reference vector for the k-th
 6 carrier according to equations (4), X_0 and Y_0 are
 7 coordinates of the reference vector, A_0 is an amplitude of
 8 the reference vector, and

9

$$B_k = A_0 * ((X_0 + dX_{rk})^2 + (Y_0 + dY_{rk})^2)^{0.5}. \quad (24c)$$

11

12 The amplitude ratio δA_k , in its turn, can be expressed
 13 through the estimate of the amplitude of the corrected
 14 reference vector. To the extent that corrected amplitude A_c
 15 is equal to

16

$$A_c = ((X_0 + dX_{rk})^2 + (Y_0 + dY_{rk})^2)^{0.5}, \quad (25)$$

18

19 then

20

$$\delta A_k = A_0 / A_c = A_0 / ((X_0 + dX_{rk})^2 + (Y_0 + dY_{rk})^2)^{0.5}. \quad (26)$$

22

23 Substituting equations (26) and (24) into equation (23),
 24 the following equalization algorithm is obtained:

$$X_{ke} = \{1 / ((X_0 + dX_{rk})^2 + (Y_0 + dY_{rk})^2)\} \{((A_0)^2 + dX_{rk}X_0 + dY_{rk}Y_0)X_k - (dX_{rk}Y_0 - dY_{rk}X_0)Y_k\} \quad (27a)$$

$$Y_{ke} = \{1 / ((X_0 + dX_{rk})^2 + (Y_0 + dY_{rk})^2)\} \{((A_0)^2 + dX_{rk}X_0 + dY_{rk}Y_0)Y_k + (dX_{rk}Y_0 - dY_{rk}X_0)X_k\}, \quad (27b)$$

Expressions (27) are a general algorithm of the dynamic equalizer, which transforms the output of the static equalizer (X_k, Y_k) into a completely equalized vector (X_{ke}, Y_{ke}) .

It will be appreciated by those skilled in the art that equations (27) can be simplified by the proper choice of the reference signal (vector). If, for example, the reference signal has coordinates $X_0=1$ and $Y_0=0$, equations (27a) and (27b) reduce to the following simple expressions:

$$X_{ke} = R_k(X_k + (dX_{rk}X_k + dY_{rk}Y_k)), \quad (28a)$$

$$Y_{ke} = R_k(Y_k + (dX_{rk}Y_k - dY_{rk}X_k)), \quad (28b)$$

where $R_k = 1 / ((1 + dX_{rk})^2 + dY_{rk}^2)$.

1 It can be seen that equations (28) differ from
2 equations (8) with respect only to the amplitude
3 coefficient R_k .

4
5 Algorithms (27) and (28) completely solve the problem
6 of per-carrier equalization, but they have appear to have
7 the disadvantage of two-step signal processing: i.e., in
8 the first step the received signal is transformed into a
9 preliminarily equalized vector (X_k, Y_k) , and in the second
10 step the preliminarily equalized vector (X_k, Y_k) is
11 transformed into a finally equalized vector (X_{ke}, Y_{ke}) .
12 Actually, in this case the static and dynamic equalizers
13 operate independently, and require double processing.

14
15 According to another aspect of the invention, the two-
16 step signal processing disadvantage is overcome by
17 combining static and dynamic equalization functions into a
18 one-step adaptive procedure. For this purpose, the static
19 equalizer algorithm will be considered in detail. In
20 particular, the static equalizer, acting during the
21 preamble, provides the receiver with equalization vector
22 (X_{kT}, Y_{kT}) for the k -th carrier. This vector does not change
23 during data transmission session. Static equalization
24 consists in multiplication of the received k -th carrier

1 vector (X_{kR}, Y_{kR}) and the equalization vector (X_{kT}, Y_{kT}) . The
 2 result of this multiplication is the equalized vector
 3 (X_k, Y_k) , having components defined by:

$$4 \quad X_k = X_{kT}X_{kR} - Y_{kT}Y_{kR}, \quad (29a)$$

$$5 \quad Y_k = X_{kT}Y_{kR} + Y_{kT}X_{kR}. \quad (29b)$$

6
 7
 8 Substituting equations (29) into (27), the full
 9 equalization algorithm is obtained which combines static
 10 (preliminary) equalization and dynamic (adaptive)
 11 equalization.

12
 13 Again, if the reference signal has coordinates $X_0=1$ and
 14 $Y_0=0$, the complete equalization algorithm reduces as
 15 follows:

$$16 \quad X_{ke} = (R_k(X_{kT} + dX_{rk}X_{kT} + dY_{rk}Y_{kT})) * X_{kR} - (R_k(Y_{kT} + dX_{rk}Y_{kT} - dY_{rk}X_{kT})) * Y_{kR},$$

$$17 \quad (30a)$$

$$18 \quad Y_{ke} = (R_k(X_{kT} + dX_{rk}X_{kT} + dY_{rk}Y_{kT})) * Y_{kR} + (R_k(Y_{kT} + dX_{rk}Y_{kT} - dY_{rk}X_{kT})) * X_{kR}.$$

$$19 \quad (30b)$$

20
 21 where X_{kR} and Y_{kR} are the quadrature components of the
 22 received, nonequalized k-th carrier signal, X_{kT} and Y_{kT} are
 23 components of the preliminary equalization vector (static
 24 vector) for the k-th carrier, dX_{rk} and dY_{rk} are estimates of

1 the differential quadrature components of the reference
 2 signal for the k-th carrier, and $R_k = 1/((1+dX_{rk})^2 + dY_{rk}^2)$ is
 3 the estimate of the amplitude correction for the k-th
 4 carrier.

5
 6 It should be appreciated that the values in the
 7 brackets of equations (30) are the corrected components of
 8 the equalization vector, and the combined static-dynamic
 9 equalization process involves the multiplication of the
 10 received k-th carrier vector (X_{kR}, Y_{kR}) and corrected
 11 equalization vector with components

$$13 \quad X_{kTc} = R_k(X_{kT} + dX_{rk}X_{kT} + dY_{rk}Y_{kT}), \quad (31a)$$

$$14 \quad Y_{kTc} = R_k(Y_{kT} + dX_{rk}Y_{kT} - dY_{rk}X_{kT}). \quad (31b)$$

15
 16
 17 It should also be noted that components of the
 18 equalization vector (31) do not require correction with the
 19 symbol rate. In other words, they may be corrected, for
 20 example, once per S symbols, where S depends on speed of
 21 change of the channel parameters. At the i-th step of
 22 equalization, the current components $X_{kT}(i)$ and $Y_{kT}(i)$ are
 23 expressed through the previous (i-1)-th components
 24 according to the following recurrent formula:

1

$$2 \quad X_{kT}(i) = R_k(X_{kT}(i-1) + dX_{rk}X_{kT}(i-1) + dY_{rk}Y_{kT}(i-1)), \quad (32a)$$

$$3 \quad Y_{kT}(i) = R_k(Y_{kT}(i-1) + dX_{rk}Y_{kT}(i-1) - dY_{rk}X_{kT}(i-1)). \quad (32b)$$

4

5 Finally, the equalization algorithm as a whole can be
6 represented using equations (30) through (32) as follows:

7

$$8 \quad X_{ke} = X_{kT}(i) * X_{kR} - Y_{kT}(i) * Y_{kR}, \quad (33a)$$

$$9 \quad Y_{ke} = X_{kT}(i) * Y_{kR} + Y_{kT}(i) * X_{kR}. \quad (33b)$$

10

11 Turning now to Figure 12, a flow-chart is seen for
12 implementing the preferred per-carrier adaptive equalizer
13 of the invention. The equalizer is adapted for
14 multicarrier wireless systems and utilizes estimates of
15 differential quadrature components of a reference vector.

16

17 As seen in Fig. 12, the carrier vectors (X_{kR}, Y_{kR}) output
18 from an FFT are equalized (corrected) at 830 by
19 multiplication with the current equalization vector (X_{kT}, Y_{kT})
20 according to

21

$$22 \quad X_{ke} = X_{kT} * X_{kR} - Y_{kT} * Y_{kR}, \quad (34a)$$

$$23 \quad Y_{ke} = X_{kT} * Y_{kR} + Y_{kT} * X_{kR} \quad (34b)$$

24

1 where X_{ke} and Y_{ke} are equalized (corrected) components of the
 2 received k-th carrier, and the transformation of equations
 3 (34) is carried out for all carriers $k=1,2, \dots, K$, where K
 4 is the number of carriers of the multicarrier signal. The
 5 current equalization vector may be a predetermined
 6 equalization vector provided from the preamble training
 7 procedure (static equalization), or may be an equalization
 8 vector estimation from the previous step of adaptive
 9 procedure (dynamic equalization).

10

11 As was described previously, at 836 the equalized
 12 signals (X_{ke}, Y_{ke}) are utilized in the multicarrier demapper
 13 for making decisions (X_{kd}, Y_{kd}) which are typically
 14 constellation points nearest to the signals (X_{ke}, Y_{ke}) and
 15 which are typically fed to a hard decoder (not shown).
 16 Along with the decisions (X_{kd}, Y_{kd}) , the demapper 836
 17 generates parameters of signal reduction such as Δ_k , A_0 and
 18 a_k (or indications thereof such as A_0/a_k) which are stored in
 19 parameters memory 838. Differential quadrature components
 20 of the equalized carriers may then be calculated at 840 as
 21 follows:

22

$$23 \quad dX_k = (X_{ke} - X_{kd}) , \quad (35a)$$

$$24 \quad dY_k = (Y_{ke} - Y_{kd}) \quad (35b)$$

1
 2 and preferably fed to a soft decoder (not shown). The
 3 differential quadrature components of the received carriers
 4 dx_k and dy_k are then reduced at 844 according to the
 5 following reduction algorithm:

6

$$7 \quad dx_{kred} = (A_0/a_k)(dx_k \cos \Delta_k - dy_k \sin \Delta_k), \quad (36a)$$

$$8 \quad dy_{kred} = (A_0/a_k)(dy_k \cos \Delta_k + dx_k \sin \Delta_k), \quad (36b)$$

9

10 where dx_{kred} and dy_{kred} are the reduced differential quadrature
 11 components of the k-th carrier, Δ_k is the phase difference
 12 between the decision and reference vectors of the k-th
 13 carrier, a_k is the amplitude of the decision vector of the
 14 k-th carrier, A_0 is the amplitude of the reference vector,
 15 and $k=1,2, \dots, K$, where K is the number of carriers of the
 16 multicarrier signal.

17

18 Sequences of reduced differential quadrature
 19 components dx_{kred} and dy_{kred} for each carrier are then averaged
 20 at 846 separately for that carrier over an interval of N
 21 symbols according to:

22

$$\begin{aligned}
 1 \quad dX_{rk} &= (1/N) \sum dX_{kred}(j) = (A_0/N) \sum_{j=1}^N (dX_k(j) \cos \Delta_k(j) - \\
 2 \quad & \quad dY_k(j) \sin \Delta_k(j)) / a_k(j), \quad (37a)
 \end{aligned}$$

$$\begin{aligned}
 3 \quad dY_{rk} &= (1/N) \sum dY_{kred}(j) = (A_0/N) \sum_{j=1}^N (dY_k(j) \cos \Delta_k(j) + \\
 4 \quad & \quad dX_k(j) \sin \Delta_k(j)) / a_k(j) \quad (37b)
 \end{aligned}$$

5
 6 where dX_{rk} and dY_{rk} are estimates of the differential
 7 quadrature components of the reference signal for the k -th
 8 carrier, $dX_k(j)$ and $dY_k(j)$ are the differential quadrature
 9 components of the k -th carrier at the j -th symbol, $\Delta_k(j)$ is
 10 the phase difference between the decision and reference
 11 vectors for the k -th carrier at the j -th symbol, and $a_k(j)$
 12 is the amplitude of the decision vector of the k -th carrier
 13 at the j -th symbol.

14
 15 Once sequences of reduced differential quadrature
 16 components are averaged for each carrier, a new
 17 equalization vector (X_{kT}, Y_{kT}) can be calculated at 850. The
 18 equalization vector depends on the reference vector, but
 19 for simplification purposes a reference vector $(1,0)$ may be
 20 chosen such that the equalization vector can be calculated
 21 as follows:

22

$$1 \quad X_{kT}(i) = R_k(X_{kT}(i-1) + dX_{rk}X_{kT}(i-1) + dY_{rk}Y_{kT}(i-1)), \quad (38a)$$

$$2 \quad Y_{kT}(i) = R_k(Y_{kT}(i-1) + dX_{rk}Y_{kT}(i-1) - dY_{rk}X_{kT}(i-1)), \quad (38b)$$

3

4 where $X_{kT}(i)$ and $Y_{kT}(i)$ are components of the new
 5 equalization vector for the k -th carrier at the current
 6 i -th step of adaptation; $X_{kT}(i-1)$ and $Y_{kT}(i-1)$ are components
 7 of the equalization vector for the k -th carrier at the
 8 previous $(i-1)$ -th step of adaptation, and $R_k = 1/((1+dX_{rk})^2 +$
 9 $dY_{rk}^2)$. The new equalization vector is then fed back to the
 10 carrier signals correction block 830 for correcting the
 11 Fourier transformed multicarrier signal according to
 12 equations (34a) and (34b).

13

14 It should be noted that the demapper and the
 15 differential components unit are not specific units of the
 16 proposed equalizer. They are typically used in any
 17 receiver. As it is indicated in Fig.12, demapper output is
 18 fed to a hard decoder, and the differential components
 19 output is fed to a soft decoder.

20

21 It will be appreciated by those skilled in the art
 22 that the flow charts of Figures 5-8 and 10-12 may be
 23 implemented in hardware, software, firmware, dedicated

1 circuitry or programmable logic, digital signal processors,
2 ASICS, or any combination of them.

3
4 There have been described and illustrated herein
5 several embodiments of a pilotless, wireless,
6 telecommunications apparatus, systems and methods. While
7 particular embodiments of the invention have been
8 described, it is not intended that the invention be limited
9 thereto, as it is intended that the invention be as broad
10 in scope as the art will allow and that the specification
11 be read likewise. Thus, with respect to all of the
12 disclosed embodiments of the invention, while particular
13 reference vectors have been disclosed, it will be
14 appreciated that other reference vectors could be utilized
15 as well. In addition, while particular mechanisms and
16 criteria for unreliable symbol exclusion have been
17 disclosed, it will be understood that other criteria and
18 mechanisms can be used. Also, while embodiments of the
19 invention have been shown in the drawings in flow-chart
20 format with particular function blocks, it will be
21 recognized that the functionality of various of the blocks
22 could be split or combined without affecting the overall
23 approach of the invention. Further, while the invention
24 was disclosed with reference to both a hard decoder and a

1 soft decoder, it will be appreciated that the receiver need
2 not include both a hard and a soft decoder, and that one or
3 the other will suffice. Thus, the current decision could
4 be sent to the soft decoder. It will therefore be
5 appreciated by those skilled in the art that yet other
6 modifications could be made to the provided invention
7 without deviating from its spirit and scope as claimed.
8